<u>**Amendments to the Specification**</u>

Please replace paragraph [0003] with the following amended paragraph:

[0003] The need for a high-quality factor (Q), low insertion loss tunable filter pervades a wide range of microwave and RF applications, in both the military, *e.g.*, RADAR, communications and electronic intelligence (ELINT), and the commercial fields such as in various communications applications, including cellular. Placing a sharply defined bandpass filter directly at the receiver antenna input will often eliminate various adverse effects resulting from strong interfering signals at frequencies near the desired signal frequency in such applications. Because of the location of the filter at the receiver antenna input, the insertion loss must be very low <u>so as</u> to not degrade the noise figure. In most filter technologies, achieving a low insertion loss requires a corresponding compromise in filter steepness or selectivity. In the present invention, the extremely low loss property of high-temperature superconductor (HTS) filter elements provides an attractive solution, achieving a very low insertion loss yet simultaneously allowing a high selectivity/steepness bandpass definition.

Please replace paragraph [0007] with the following amended paragraph:

[0007] In addition to the intermodulation/dynamic range problems of varactors, these conventional tuning devices also have serious limitations in Q, or tuning selectivity. Because the varactors operate by varying the depletion region width of a P-N junction, this means that at lower reverse biases (higher capacitances), there is a substantial amount of undepleted moderately-doped semiconductor material between the contacts and the junction that offers significant series resistance (R_{ac}) to ac current flow. Since the Q of a varactor of junction capacitance C_j and series resistance R_{ac} at the signal frequency f is given by $Q = 1/(2\ f\ C_j\ R_{ac})$,

this means that the varactor Q values are limited, particularly at higher frequencies. For example, a typical commercial varactor might have C_j = 2.35 pF with R_{ac} =1.0 Ω at V_r =-4V, or C_j = 1.70pF with R_{ac} =0.82 Ω at V_r =-10V, corresponding to Q values at f = 1.0 GHz of Q = 68 at V_r = -4V or Q= 114 at V_r = -10V (or f = 10.0 GHz values of Q = 6.8 and Q = 11.4, respectively). Considering that an interesting X-band (f = 10 GHz) RADAR application might want a bandwidth of ~~f~~= 20MHz for the full width at half-maximum (FWHM), corresponding to a Q = 500 quality factor, we see that available varactors have inadequate Q (too much loss) to meet such requirements. While the mechanisms are different, this will very likely apply to the use of ferroelectrics or other "tunable materials." A general characteristic of materials which exhibit the field-dependent dielectric constant nonlinearities (that makes them tunable) is that they exhibit substantial values of the imaginary part of the dielectric constant (or equivalently, loss tangent). This makes it unlikely that, as in varactors, these "tunable materials" will be capable of achieving high Q's, particularly at high signal frequencies.

Please replace paragraph [0036] with the following amended paragraph:

[0036] In this floating plate HTS variable capacitor tunable filter structure 25, wide HTS lines (potentially much wider than illustrated in the top view of Figure 1b) may be used to define an inductor 40 (or alternatively, a transmission line segment) coupled at full width between the fixed plates 10 and 15 to achieve very high Q values. The inductor 40 may have an electrical length substantially equal to a one-half wavelength corresponding to a fundamental resonant frequency so that it becomes a simple L-C resonator circuit. Alternatively, the inductor 40 may have an electrical length substantially equal to multiples of this one-half wavelength. Moreover, the inductor 40 may be substantially shorter than one-half wavelength, thus approximating a

lumped circuit element. The resulting equivalent circuit is portrayed in Fig 1c. Two variable capacitors 60 and 65 are in series with the inductor 40. The mechanical driver 45 of Figure 1a varies the capacitance of capacitors 60 and 65.

Please replace paragraph [0037] with the following amended paragraph:

[0037] The resulting equivalent circuit as portrayed in Fig 1c has important consequences resulting from its balanced structure. Note that a commonly used half-wavelength resonator comprises an open circuited length of microstrip transmission line as illustrated in Figure 2a. The transmission line has a length (denoted as "d") substantially equal to one-half wavelength $(\frac{\lambda_0}{2})$ corresponding to the fundamental resonant frequency f_0. Such resonators are frequently tuned by use of a metal tuning screw through the cover plate near one end of the resonator to add capacitance between the end of the resonator and ground. Instead of a screw, one could use a variable capacitor C_t as disclosed by the present invention. For frequencies at or near resonance, the impedance Z_{in} seen looking into the end of the resonator matches that of a parallel tuned circuit as illustrated in 2b. For a given capacitance variation ΔC_t, there will be some resulting resonant frequency variation Δf_0.

Please replace paragraph [0038] with the following amended paragraph:

[0038] Should the half-wavelength $(\frac{\lambda_0}{2})$ resonator of Figure 2a be bent into a "hairpin" configuration as shown by the inductor 40 in Figure 1b and C_t be connected across the open ends of the resonator, the balanced structure of Figure 2c is produced. At resonance, the potentials on the two open ends of this "hairpin" resonator have opposite potential, so that the terminals of the

variable capacitor are now driven in a balanced manner. Because of the balanced operation, for whatever potential that exists at a point on the right side of the hairpin, there is an equal and opposite potential at the corresponding point on the left side of the resonator. Therefore, a virtual ground will run through the centerline <u>CL</u> of the resonator as shown in Figure 2d. It is then convenient to model the capacitance in the equivalent form wherein the single capacitor C_t is replaced by two capacitors each having twice the capacitance of the original capacitor C_t. If the circuit illustrated in Figure 2d were straightened out as in Figure 2a, there would be a capacitor of size 2Ct at both ends of the transmission line. Thus, the capacitor Ct in Figure 2c is four times more effective for tuning than it is in the structure illustrated in Figure 2a.

Please replaced paragraph [0040] with the following amended paragraph:

[0040] A number of alternate geometries may be employed to construct the variable capacitor used in the present invention. For example, rather than employing the preferred "split plate" design illustrated in Figure 1a, the first and second fixed plates 10 and 15 could be replaced with a single fixed plate (not illustrated). In such a design, however, both plates on either side of the capacitor gap 50 would require a coupling to the filter signal. In addition to alternate geometries, the floating plate 30 or the first and second fixed plates 10 and 15 may be alternatively constructed out of <u>a normal state conductor such as</u> gold or a similar low-loss metal rather than from HTS material. Such metals do not produce, however, Q values as high as that can be achieved using an all HTS construction.

Please replace paragraph [0043] with the following replacement paragraph:

[0043] The basic structure of this embodiment of the variable capacitor HTS tunable filter 25 is similar to that already discussed with respect to Figs. 1a and 1b. As best seen in Figure 3a, to accommodate the two folded piezoelectric driver structures 70 and to minimize parasitic capacitive effects between the variable capacitor HTS tunable filter 25 and the piezoelectric driver structures 70, the movable substrate has lateral extensions 80 and 85 which extend laterally substantially past fixed plates 10 and 15, respectively. The first fixed plate 10 and the second plate 15 are patterned from HTS material and grown on the fixed substrate 20 comprising, for example, MgO. An HTS material inductor 40 (not illustrated) is coupled between the first and second fixed plates 10 and 15. As previously described, this inductor 40 preferably has an electrical length chosen such that the inductor 40 behaves as a simple L-C resonator circuit. As illustrated in Fig 3b, the first and second fixed plates 10 and 15 each has a lead 16 so that a filter signal may be coupled to the variable capacitor HTS tunable filter 25 (seen in Figure 3a). The variable capacitor HTS tunable filter structure 25 is completed by the addition of a floating plate 30 (drawn transparent in the plan view in Figure 3b) patterned from thin-film HTS material grown on the movable substrate 35 comprising, for example, MgO. As seen in Figure 3a, a pair of folded piezoelectric drivers 70 is coupled to the lateral extensions 80 and 85 as follows. A first piezoelectric driver 100 is attached, using for example, low-temperature indium alloy solder (which also may be used for the remaining required piezoelectric attachments), to the fixed substrate 20 at a location adjacent to the lateral extension 80 of movable substrate 35. The first piezoelectric driver 100 is attached at its opposite end to a coupling member 110. Also attached to coupling member 110 is the second piezoelectric driver 115. Coupling member 110 may be constructed from a suitable material such as MgO. Alternatively, coupling member 110 may simply be a low temperature indium solder bond or the

like in order to couple the first and second piezoelectric drivers 100 and 115 together. The second piezoelectric driver 115 is attached at its opposite end to the lateral extension 80 of movable substrate 35, forming one of the pair of folded piezoelectric driver structures 70 wherein the first piezoelectric driver 100 and the second piezoelectric driver 115 are substantially parallel to one another. In this fashion, when the first piezoelectric driver 100 shortens and the second piezoelectric driver 115 lengthens, the gap 50 between the capacitor plates is shortened, affecting the tunable filter 25 frequency response. The other folded piezoelectric driver structure 70 is coupled to the lateral extension 85 of movable substrate 35 in an analogous fashion as follows. A third piezoelectric driver 120 is attached to lateral extension 85 at one end and attached to a coupling member 125 at its opposite end. Also attached to coupling member 125 is a fourth piezoelectric driver 130. The fourth piezoelectric driver 130 is attached at its opposite end to the fixed substrate 20 at a location adjacent to lateral extension 85 so that the third piezoelectric driver 120 and the fourth piezoelectric driver 130 are substantially parallel to one another, thereby forming the other piezoelectric driver structure 70. To suppress bending vibrations, an optional bridge 140, comprised of, *e.g.*, MgO, may span between the coupling members 110 and 120.

Please replace paragraph [0044] with the following amended paragraph:

[0044] In one embodiment, each piezoelectric driver is coupled to a tuning signal having a tuning signal voltage, Vt through electrodes 105. To implement the folded piezoelectric actuator structure 70, a proper combination of tuning signal voltage (Vt) polarity and the orientation of the piezoelectric material ("poling" direction) is necessary. For example, in one embodiment, with positive voltage ±Vt, the first and fourth piezoelectric drivers 100 and 130 get

shorter and the second and third piezoelectric drivers 115 and 120 lengthen, driving the floating plate 30 closer toward the fixed plates 10 and 15, decreasing the gap 50 and increasing the capacitance. Analogously, in the same embodiment, a negative voltage -Vt would lengthen the first and fourth piezoelectric drivers 100 and 130 and shorten the second and third piezoelectric drivers 115 and 120. Thus, with a negative voltage -Vt, the gap 50 between the plates increases, decreasing the capacitance. The mode of piezoelectric material (typically lead zirconate titanate, "PZT") operation is as illustrated at the lower left in Figure 3a, in which a bias voltage, Vt, applied across the thin "Z" dimension of a plate causes either shortening in "X" and lengthening in "Y," or lengthening in "X" and shortening in "Y," depending on the polarity of voltage Vt. In such an embodiment, the orientation of the paired piezoelectric drivers with their respective electrodes within each folded piezoelectric driver structure may be denoted as "opposite" to each other.

Please replace paragraph [0046] with the following amended paragraph:

[0046] The HTS variable capacitor structure requires, at least in the vertical dimension, precision mechanical fabrication and assembly. The precision fabrication is expressed principally as a flatness and peak non-planarity (bump height) specification on the facing surfaces of the fixed plates 10 and 15 and floating plate 30. To avoid short circuiting between opposing capacitor plates as the mean separation distance between the surfaces is reduced to the micron range (or even submicron range), there must not be any substantial protrusions from the facing surfaces of the plates. In one embodiment, the as-grown HTS surfaces are scanned for the presence of nodules or other "bumps" protruding from the surface and laser ablation or other techniques are employed to remove any that are too high. Note that in a capacitor plate, there is

no significant performance penalty for having a few holes in the conductor film, so such a post-growth "re-figuring" of the surfaces need not be planarizing, just effective in removing "bumps" ("dips" are not a problem). The critical problem in precision assembly is establishing the correct initial (Vt = 0) height (separation) and parallelism of the capacitor plates, preferably to a fraction of a micron (at least in parallelism). In one embodiment, this precision vertical alignment may be achieved by means of a self-alignment process based on an art-recognized type of removable deposited spacer technique. For example, if a 5 μm nominal spacing were desired, then a uniform 5 μm (post-develop) thick layer of photoresist could be spun onto the previously-patterned HTS fixed substrate 20 (shown as MgO in Figure ~~2~~ 1a) and exposed and developed to leave, for example, six "pucks," 5μm high (one at each corner and two along the long sides of where the movable substrate 35 will be placed. The movable substrate 35 may then be aligned and pressed down onto these posts while the piezoelectric drivers 70 are attached on each end with low-temperature (*e.g.*, indium alloy) solder. After assembly is complete, the posts are dissolved away in a suitable solvent to free the floating plate 30 for unimpeded vertical motion when tuning bias voltage, Vt, is applied. (Note that the pucks may be made of photosensitive material, or other, more easily dissolved material (like poly-methyl methacrylate (PMMA)) which has been patterned using photoresist.)

Please replace paragraph [0047] with the following amended paragraph:

[0047] Note in Figure 3a that the thinned (for low mass when the fastest tuning rate is desired) movable substrate 35 extends considerably beyond the ends of the HTS floating plate 30 itself, forming lateral extensions 80 and 85. The reason for these substantial gaps between the floating plate 30 and the attachment points for the second and third piezoelectric drivers 115 and

120 is to minimize stray capacitance between the HTS floating plate 30 and the folded PZT driver structures 70 to avoid energy loss/Q degradation effects. Note that this piezoelectric actuator approach provides the ideal of a tuning input (the Vt electrodes on the PZT drivers) that is completely separate from the signal path; leaving a good clearance between the floating plate 30 and the PZT drivers just insures that maximum Q is realized.

Please replace paragraph [0048] with the following amended paragraph:

[0048] To better illustrate what even a very simple tunable HTS filter of the type illustrated in Figures 1a, 1b, 3a, and 3b could do in a system application, an example of a simple, single-resonator, narrowband bandpass preselector filter for an X-band RADAR was considered. Those of ordinary skill in the art will appreciate that such an example is merely illustrative of the invention and that the example is easily scaled to illustrate operation at lower or higher frequencies. The $L_2 = 0.5384$nH, $C_2 = 0.50$ pF (for f_0=9.70GHz) parallel "tank" resonator (the dimensions and projected unloaded Q of which were discussed above) was considered with simple inductive loop coupling to a 50 ohm source and a 50 ohm load was considered. Analysis of the circuit was performed both analytically and using simulated program with integrated circuit emphasis (SPICE) (the two approaches give the same result). Since the unloaded Q of the resonator (approximately 50,000 as calculated previously) is so much higher than the intended loaded Q (a bandwidth of 20MHz [full width at half maximum (FWHM)] at $f_0 = 10$ GHz would be appropriate for a RADAR application, corresponding to a loaded $Q = f_0/ \Delta f = 500$), the ac resistance R_{ac} of the L_2-C_2 resonator was ignored. The bandwidth of this simple loop-coupled single resonator tunable bandpass filter is determined by the inductance of the input and output coupling loops (L_1 and L_3, both taken as 20% of L_2) and their coupling coefficients (K_{12} and K_{23},

both taken as 0.09) to the resonator inductor, L_2. These particular values were selected to give a -3dB bandwidth of 20MHz. Note that in the physical construction of such an HTS filter, it is not necessary that these input and output coupling loops be made of HTS material, or even lie on the cooled HTS substrate. It would be easier to have them lie well above and below the resonator substrate ~~(~~to minimize direct coupling between them, and to have their positions mechanically adjustable when the filter is setup to select the exact bandwidth desired. Note that placing these input and output loops out of contact with the cooled HTS substrate avoids the thermal conduction load associated with metallic conductors to the substrate. (This advantage would apply equally well to capacitive probe coupling to the resonator, which is another satisfactory means of making a resonator into a bandpass filter).

Please replace paragraph [0050] with the following amended paragraph:

[0050] ~~From an RF/μW performance standpoint~~ For radio frequency and microwave applications ~~(unloaded Q, tuning range, etc.)~~ the present invention offers outstanding performance (unloaded Q, tuning range, etc.) There is some question of whether the tuning speed, using only conventional PZT materials, can be brought down to the 5 μs to 10 μs range that would be desirable for frequency-hopping applications (unless operation with very small gaps proves feasible, although the use of higher performance ceramic piezoelectric or electroactive polymer materials should make this possible). However, the filters themselves are quite small, and so there would be little penalty to putting two of these filters in a cryocooler vacuum dewar and "ping-ponging" back and forth between them on successive hops. Thus, with a hopping residence time of a few milliseconds, there would be plenty of time to tune the "off duty" filter to the next hop frequency. In fact, even if undesirable piezoelectric driver hysteresis

or drift characteristics prevent obtaining precise "first-shot" open-loop tuning, there should be ample time to check the new set frequency of the "off –duty" filter and correct it if necessary. Of course, in a preferred embodiment, the present invention will use the finest available "atomic force microscopy (AFM) quality" piezoelectric actuator technology to achieve fast, open-loop operation in the simplest way possible.